Prior Art

DESCRIPTION

WIRELESS TERMINAL

The present invention relates to a wireless terminal, for example a mobile phone handset.

Wireless terminals, such as mobile phone handsets, typically incorporate either an external antenna, such as a normal mode helix or meander line antenna, or an internal antenna, such as a Planar Inverted-F Antenna (PIFA) or similar.

Such antennas are small (relative to a wavelength) and therefore, owing to the fundamental limits of small antennas, narrowband. However, cellular radio communication systems typically have a fractional bandwidth of 10% or more. To achieve such a bandwidth from a PIFA for example requires a considerable volume, there being a direct relationship between the bandwidth of a patch antenna and its volume, but such a volume is not readily available with the current trends towards small handsets. Hence, because of the limits referred to above, it is not feasible to achieve efficient wideband radiation from small antennas in present-day wireless terminals.

A further problem with known antenna arrangements for wireless terminals is that they are generally unbalanced, and therefore couple strongly to the terminal case. As a result a significant amount of radiation emanates from the terminal itself rather than the antenna.

An object of the present invention is to provide a wireless terminal having efficient radiation properties over a wide bandwidth.

According to the present invention there is provided a wireless terminal comprising a ground conductor and a transceiver coupled to an antenna feed, wherein the antenna feed is coupled to the ground conductor.

The present invention is based upon the recognition, not present in the prior art, that the impedances of an antenna and a wireless handset are similar to those of an asymmetric dipole, which are separable, and on the further

recognition that the antenna impedance can be replaced with a non-radiating coupling element.

Embodiments of the present invention will now be described, by way of example, with reference to the accompanying drawings, wherein:

Figure 1 shows a model of an asymmetrical dipole antenna, representing the combination of an antenna and a wireless terminal;

Figure 2 is a graph demonstrating the separability of the components of the impedance of an asymmetrical dipole;

Figure 3 is an equivalent circuit of the combination of a handset and an antenna;

Figure 4 is an equivalent circuit of a capacitively back-coupled handset;

Figure 5 is a perspective view of a basic capacitively back-coupled handset;

Figure 6 is a graph of simulated return loss S_{11} in dB against frequency f in MHz for the handset of Figure 5;

Figure 7 is a Smith chart showing the simulated impedance of the handset of Figure 5 over the frequency range 1000 to 2800MHz;

Figure 8 is a graph showing the simulated resistance of the handset of Figure 5;

Figure 9 is a perspective view of a narrow capacitively back-coupled handset;

Figure 10 is a graph showing the simulated resistance of the handset of Figure 9;

Figure 11 is a perspective view of a slotted capacitively back-coupled handset;

Figure 12 is a graph of simulated return loss S_{11} in dB against frequency f in MHz for the handset of Figure 11;

Figure 13 is a Smith chart showing the simulated impedance of the handset of Figure 11 over the frequency range 1000 to 2800MHz;

Figure 14 is a plan view of a capacitively back-coupled test piece;

Figure 15 is a graph of measured return loss S_{11} in dB against frequency f in MHz for the test piece of Figure 14;

Figure 16 is a Smith chart showing the measured impedance of the test piece of Figure 14 over the frequency range 800 to 3000MHz;

Figure 17 is a plan view of a capacitively back-coupled test piece using an inductive element;

Figure 18 is a graph of measured return loss S_{11} in dB against frequency f in MHz for the test piece of Figure 17; and

Figure 19 is a Smith chart showing the measured impedance of the test piece of Figure 17 over the frequency range 800 to 3000MHz.

In the drawings the same reference numerals have been used to indicate corresponding features.

Figure 1 shows a model of the impedance seen by a transceiver, in transmit mode, in a wireless handset at its antenna feed point. The impedance is modelled as an asymmetrical dipole, where the first arm 102 represents the impedance of the antenna and the second arm 104 the impedance of the handset, both arms being driven by a source 106. As shown in the figure, the impedance of such an arrangement is substantially equivalent to the sum of the impedance of each arm 102,104 driven separately against a virtual ground 108. The model could equally well be used for reception by replacing the source 106 by an impedance representing that of the transceiver, although this is rather more difficult to simulate.

The validity of this model was checked by simulations using the well-known NEC (Numerical Electromagnetics Code) with the first arm 102 having a length of 40mm and a diameter of 1mm and the second arm 104 having a length of 80mm and a diameter of 1mm. Figure 2 shows the results for the real and imaginary parts of the impedance (R+jX) of the combined arrangement (Ref R and Ref X) together with results obtained by simulating the impedances separately and summing the result. It can be seen that the results of the simulations are quite close. The only significant deviation is in the region of half-wave resonance, when the impedance is difficult to simulate accurately.

An equivalent circuit for the combination of an antenna and a handset, as seen from the antenna feed point, is shown in Figure 3. R_1 and jX_1 represent the impedance of the antenna, while R_2 and jX_2 represent the impedance of the handset. From this equivalent circuit it can be deduced that the ratio of power radiated by the antenna, P_1, and the handset, P_2, is given by

$$\frac{P_1}{P_2} = \frac{R_1}{R_2}$$

If the size of the antenna is reduced, its radiation resistance R_1 will also reduce. If the antenna becomes infinitesimally small its radiation resistance R_1 will fall to zero and all of the radiation will come from the handset. This situation can be made beneficial if the handset impedance is suitable for the source 106 driving it and if the capacitive reactance of the infinitesimal antenna can be minimised by increasing the capacitive back-coupling to the handset.

With these modifications, the equivalent circuit is modified to that shown in Figure 4. The antenna has therefore been replaced with a physically very small back-coupling capacitor, designed to have a large capacitance for maximum coupling and minimum reactance. The residual reactance of the back-coupling capacitor can be tuned out with a simple matching circuit. By correct design of the handset, the resulting bandwidth can be much greater than with a conventional antenna and handset combination, because the handset acts as a low Q radiating element (simulations show that a typical Q is around 1), whereas conventional antennas typically have a Q of around 50.

A basic embodiment of a capacitively back-coupled handset is shown in Figure 5. A handset 502 has dimensions of 10×40×100mm, typical of modern cellular handsets. A parallel plate capacitor 504, having dimensions 2×10×10mm, is formed by mounting a 10×10mm plate 506 2mm above the top edge 508 of the handset 502, in the position normally occupied by a much larger antenna. The resultant capacitance is about 0.5pF, representing a compromise between capacitance (which would be increased by reducing the separation of the handset 502 and plate 504) and coupling effectiveness (which depends on the separation of the handset 502 and plate 504). The

capacitor is fed via a support 510, which is insulated from the handset case 502.

The return loss S_{11} of this embodiment after matching was simulated using the High Frequency Structure Simulator (HFSS), available from Ansoft Corporation, with the results shown in Figure 6 for frequencies f between 1000 and 2800MHz. A conventional two inductor "L" network was used to match at 1900MHz. The resultant bandwidth at 7dB return loss (corresponding to approximately 90% of input power radiated) is approximately 60MHz, or 3%, which is useful but not as large as was required. A Smith chart illustrating the simulated impedance of this embodiment over the same frequency range is shown in Figure 7.

The low bandwidth is because the handset 502 presents an impedance of approximately 3−j90Ω at 1900MHz. Figure 8 shows the resistance variation, over the same frequency range as before, simulated using HFSS. This can be improved by redesigning the case to increase the resistance.

One way in which this can be done is to reduce the width of the handset 502, since the resistance will increase in much the same way as that of a dipole when its radius is decreased. Figure 9 shows a second embodiment having a narrow capacitively back-coupled handset 902. The handset 902 has dimensions of 10×10×100mm, while the dimensions of the capacitor 504, formed from the plate 506 and top surface 908 of the handset 902, and the support 510 are unchanged from the previous embodiment. Simulations were again performed to determine the resistance variation of this embodiment, with the results shown in Figure 10. This clearly demonstrates that use of a narrow handset provides a wider bandwidth where the resistance is higher than that of the basic configuration. The length of the handset could be optimised to give a wide bandwidth centred on a particular frequency, by shifting the resonant frequencies of the structure. For a fixed length handset, a horizontal slot (i.e. a slot across the width of the handset) could be used for the purpose of electrically shortening or lengthening the handset.

An alternative way of increasing the resistance of the case is the insertion of a vertical slot (i.e. a slot parallel to the length, or major axis, of the

handset). Figure 11 shows a third embodiment having a slotted capacitively back-coupled handset 1102, with a 33mm deep slot 1112 in the case, together with a capacitor 504. The dimensions of the capacitor 504, formed from the plate 506 and top surface 1108 of the handset 1102, and the support 510 are unchanged from the previous embodiments. The presence of the slot 1112 significantly increases the resistance of the case, as seen by the transceiver, in the region of 1900MHz, allowing the low-Q case to be matched to 50Ω without a significant loss of bandwidth.

The return loss S_{11} of this embodiment was again simulated using HFSS, with the results shown in Figure 12 for frequencies f between 1000 and 2800MHz, using a similar two inductor matching network to that used for the basic embodiment. The resultant bandwidth at 7dB return loss is greatly improved at approximately 350MHz, or 18%, which is approaching that required to cover UMTS and DCS 1800 bands simultaneously. A Smith chart illustrating the simulated impedance of this embodiment over the same frequency range is shown in Figure 13.

A test piece was produced to verify the practical application of the simulation results presented above. Figure 14 is a plan view of the test piece, which comprises a copper ground plane 1402 having dimensions 40×100mm on a 0.8mm thick FR4 circuit board (with a measured dielectric constant of 4.1). A 3×29.5mm slot 1412 is provided in the ground plane and a 10×10mm plate 506 is located 2mm above the corner of the ground plane 1402. The plate and co-extensive portion of the ground plane 1402 form a parallel plate capacitor, as in the embodiments described above. The capacitor is fed via a co-axial cable 1404 attached to the rear surface of the circuit board and a vertical pin 510.

The return loss S_{11} of this embodiment was measured without matching, which was then added in simulations. The matching added was a 3.5nH series inductor and a 4nH shunt inductor, similar to that used in the simulations described above. Results are shown in Figure 15 for frequencies f between 800 and 3000MHz. The resultant bandwidth at 7dB return loss is approximately 350MHz centred at 1600MHz, or 22%, which is approximately

the fractional bandwidth required to cover UMTS and DCS 1800 bands simultaneously. A Smith chart illustrating the impedance of this embodiment over the same frequency range is shown in Figure 16.

The embodiments disclosed above are based on capacitive coupling. However, any other sacrificial (non-radiating) coupling element could be used instead, for example inductive coupling. Also, the coupling element could be altered in order to aid impedance matching. For example, capacitive coupling could be achieved via an inductive element which has the advantage of requiring no further matching components.

As an example of this latter technique a further test piece was produced, illustrated in plan view in Figure 17. This piece is similar to that shown in Figure 14, with the difference that the plate 506 is slightly offset from the corner of the ground plane 1402 and is no longer completely metallised: instead a spiral track 1706 is provided, connected at one end to the feed pin 501. The length of the track 1706 is chosen to provide resonance at the required frequency, approximately 1600MHz in this embodiment. The track 1706 is fed via a stripline 1704 on the rear surface of the circuit board.

The return loss S_{11} of this embodiment was measured without matching. Results are shown in Figure 18 for frequencies f between 800 and 3000MHz. The resultant bandwidth at 7dB return loss is approximately 135MHz centred at 1580MHz, or 9%, and it is believed that this bandwidth could be improved significantly by further optimisation and matching. A Smith chart illustrating the impedance of this embodiment over the same frequency range is shown in Figure 19.

In the above embodiments a conducting handset case has been the radiating element. However, other ground conductors in a wireless terminal could perform a similar function. Examples include conductors used for EMC shielding and an area of Printed Circuit Board (PCB) metallisation, for example a ground plane.

From reading the present disclosure, other modifications will be apparent to persons skilled in the art. Such modifications may involve other features which are already known in the design, manufacture and use of

wireless terminals and component parts thereof, and which may be used instead of or in addition to features already described herein. Although claims have been formulated in this application to particular combinations of features, it should be understood that the scope of the disclosure of the present application also includes any novel feature or any novel combination of features disclosed herein either explicitly or implicitly or any generalisation thereof, whether or not it relates to the same invention as presently claimed in any claim and whether or not it mitigates any or all of the same technical problems as does the present invention. The applicants hereby give notice that new claims may be formulated to such features and/or combinations of features during the prosecution of the present application or of any further application derived therefrom.

In the present specification and claims the word “a” or “an” preceding an element does not exclude the presence of a plurality of such elements. Further, the word “comprising” does not exclude the presence of other elements or steps than those listed.